N-path filters and Mixer-First Receivers: A Review

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Abstract—To realize a Software Defined Radio covering the mainstream 0.5-6 GHz wireless communication bands, new SAW-less radio receiver architectures are being explored which realize selectivity in a more flexible and programmable fashion. N-path filters and mixer-first receivers can offer high-linearity high-Q RF-filtering around a center frequency defined by a digital clock, which offers the desired flexible programmability. This paper reviews recent research on N-path filters and mixer-first receivers, identifies advances in performance analysis, circuit performance and applications.

Keywords—Mixer-first receiver, N-path filter, Frequency Translated filtering, passive mixer, CMOS, switch-R-C circuit, software defined radio, reconfigurable radio, RF, wireless

I. INTRODUCTION

In a world with ever more mutually interfering wireless devices, selectivity by high-Q filtering is of paramount importance. Traditionally, this filtering is implemented by high-Q LC-filters or devices that can be modelled as such (e.g. devices exploiting transmission lines, Surface Acoustic Waves (SAW) or Bulk Acoustic Waves (BAW)). Although these devices can achieve very good selectivity and excellent linearity, they lack the programmability wanted for software defined radio. Moreover, compared to solutions exploiting CMOS IC technology, size and cost are key issues.

CMOS IC technology is very suitable to realize well-matched highly linear oxide-capacitors, while Moore's law also offers faster digital circuits and better switches in terms of low on-resistance and parasitic capacitance. As N-path filters exploit switches and capacitors, see e.g. Fig. 1, while offering high-Q filtering around a digitally programmable switching frequency, this explains the recent interest in N-path filters and mixer-first receivers for software defined radio.

Historically, different names have been used like "Narrow Band-Pass filter Using Modulation" [1], "Commutated Network" [2], "Comb Filter using Commutated Capacitors" [2, 3], "Sampled-Data Band-Pass filter" [4], Frequency-Translated Filter [4], "N-Path filter" [5]. We will use the generalized term N-path filter here, as in the filter survey by Moschytz [6]: "N identical networks $H(j\omega)$ are cyclically switched into the signal path. Thus, the total N-phase impulse network has time-variable characteristics. If a low-pass filter is used to provide the $H(j\omega)$ function in each of the paths, the cyclical switching process causes a low-pass to bandpass transformation." Note that we do not assume a uni-lateral transfer for each of the N signal paths, although this is sometimes convenient to simplify analysis, e.g. as in [5].

Fig. 1. Commutated band-pass and band-stop filter and their frequency response as predicted by Smith in 1953 [2]. Early commutators were implemented as motor driven rotating mechanical switches.

This paper reviews the literature on N-path filters and mixer-first receivers, trying to identify advances in performance analysis (section II), steps in the evolution of mixer-first receivers (section III) and N-path filters (section IV), concepts to lower power dissipation (section V) and new applications (section V). The primary focus will be on high linearity passive-first frequency-selective switch-R-C circuits. Active feedback techniques as reviewed in [7] and wideband approaches like in [8] are not covered, but can be found in other reviews like [9-11]. The reference list focusses on journal papers, even if earlier conference papers exist.

II. PERFORMANCE ANALYSIS OF N-PATH FILTERS

Although the switch-R-C circuits in Fig. 1 look simple, mathematical analysis of their transfer function is complicated by their time-variant nature combined with the memory effect of the capacitors. We will describe different simplifications that have been used over time to make performance analysis tractable.

Early papers [2, 3, 5] make simplifying assumptions ultimately leading to a kind of "sampled data" approximation, hence the title "Sampled Data Band-Pass Filter" [4]. Assuming such ideal sampling there is no loss and this leads to an interesting analogy of the commutator in Fig. 1 with a transmission line: if the rotator cycle-time is equal to the period of a sinewave input signal, the voltages on the capacitors become samples of one period of the sinewave.

Fig. 2. 4-path switch-R-C filter: a) time-domain waveforms; b) Simulated transfer function (dashed line) and approximate R-L-C model (drawn) [12] ($R=50\Omega$, $C=50pF$, 1Ω switch resistance, 1 GHz clock with 25% duty cycle).

After one full commutator cycle, the new incoming sinusoid is hence “confronted with” a sample stored on the capacitor one period ago. This behavior resembles that of an open transmission line of length $T/2$, which fully reflects the input signal of one period ago [2, 3]. As a result, the transfer function shown in Fig. 1 has the well-known comb filter behavior of transmission line filters, with identical gain for all harmonics. Although the identical gain at harmonics is a somewhat crude approximation (see Fig. 2 for $N=4$ [12]), it gives useful first order insight. Moreover, [2, 3] already rather accurately estimated the -3dB bandwidth of the filter:

$$f_{-3dB} = \frac{1}{\pi N R C} \quad (1)$$

This means that the bandwidth is inversely proportional to the *total capacitance* NC . Essentially the N-path filter down-converts the RF-signal, low-pass filters it, and then up-converts it again to RF (the switches in Fig. 1 serve both the down- and up-conversion purpose). Hence the low-pass filter characteristic is simply shifted around the commutator-frequency. Intuitively, the RF-bandwidth is twice that of an RC-lowpass filter, multiplied by N as each capacitor “sees” the resistor R only for a duty cycle $1/N$.

From this simple analysis we can gain some very relevant insights explaining the interest in N-path filters for software defined radio: 1) the center frequency of the filter is accurate and widely programmable using digital clocking; 2) the RF-bandpass roll-off is essentially the same as for a real-pole RC low-pass filter, but frequency shifted; 3) the equivalent $Q = f_s/f_{-3dB}$ can approach SAW-filter Q-values, as GHz switching frequencies are feasible in nanometer CMOS, while f_{-3dB} can be in the low MHz region if we add sufficient capacitance; 4)

Fig. 3. 4-Path mixer-first receiver with 25% duty-cycle clocks, differential RF-input V_{in} , low-pass baseband quadratures outputs $V_{I+}, V_{I-}, V_{Q+}, V_{Q-}$ and band-pass filtered RF-output V_{out} . Even harmonic responses cancel. Switch resistance limits out-of-band filter rejection, as modelled via R_s and $2R_{sw}$.

linearity perspectives are good as MOSFET passive mixers can have very good linearity.

The early N-path filters use the output signal in a time-continuous fashion. In the end of the 60’s research attention gradually shifted to time discrete signal processing, where the output signal is only interpreted at certain instants. To achieve a sharper filter roll-off, complex pole pairs are wanted. This was implemented combining switched capacitor networks first with finite gain amplifiers [13] and later with high-gain Op-amps [6]. Although N-path filters with time-continuous output are still mentioned in the filter survey of [6] and in 70s textbooks on switched-capacitor filters, literature references fades out in the 80s. In papers on N-path filters that we did find, the filter output is interpreted in a *time discrete manner*, as is apparent from the z-domain analysis that is used [14]. Only in the last decade N-path filters with a time-continuous output regained research interest [15-25]. Moreover, mixer-first receivers exploiting N-path frequency translated filtering were proposed [26-32]. Fig. 3 shows a mixer with differential RF-input exploiting two sets of switches driven by a 25% duty cycle. Capacitor baseband voltages $V_{I+}, V_{I-}, V_{Q+}, V_{Q-}$ are now used as output instead of the RF voltage V_{out} . Note however that RF filtering still takes place, as the switches not only down-convert to baseband, but also up-convert so that N-path filtering occurs at the RF-input.

This review focusses on N-path switch-R-C circuits with a time-continuous output. In contrast to circuits with a time discrete output that use sampling with fast settling, time-continuous N-path filters with good frequency selectivity show extremely slow settling, analogous to high-Q passive filters. We can actually define two regions for a simple switch-R-C circuit with on-time T_{on} [16, 33, 34]:

$$\text{Sampling region: } T_{on} \gg 2RC \quad (2a)$$

$$\text{Mixing region: } T_{on} \ll 2RC \quad (2b)$$

Using a simplified analysis based on energy conservation while assuming very big RC times, Cook et al [26] found .375

Fig. 4. Constant Noise Figure contours for a switch-R-C down-conversion mixer for arbitrary RC-time, switching frequency and duty cycle. Note the lower noise figure in the mixing-region for big RC due to less noise folding.

as best achievable noise figure for a 4-path mixer with 25% duty cycle in each path.

This result was generalized and confirmed by the unified analysis of samplers and mixers in [16]. In Fig. 4 [35], noise figure of a switch-R-C circuit is plotted for arbitrary values of switching frequency, RC-time and duty cycle. The theoretical noise figure for a 4-path mixer is 6.9dB. When combining 4 output signals, SNR is improved 4x (6dB), leading to a theoretical minimum noise figure of 0.9dB, fitting to the $\pi^2/8$ [26]. This demonstrates the low-noise perspective of N-path filters and mixers.

Using the analysis procedure in [16], N-path filter transfer has been analyzed in [12] and the equivalent R-L-C filter model of Fig. 3 was derived for arbitrary N to fit the fundamental response. The analysis was also used to find the transfer function for N-path notch filters [23], gain-boosted N-path filters [36], and an N-path filter preceded by a series-inductor to realize time-continuous low-pass pre-filtering [37]. Such pre-filtering is important, as N-path filters show folding of RF-signals around some harmonics of the clock-frequency. The first folding frequency band is around $(N-1)f_s$, similar to aliasing in discrete time circuits, albeit with more attenuation. Using more paths, e.g. 8 instead of 4, allows for cancelling more folding products. The analysis of [16, 17, 23] is suitable to analyze folding, as no assumptions about the input frequency are made. Whereas this unified analysis of mixing and sampling (see Fig. 4) analyzes the continuous time output, a recent unified analysis [33, 34] focusses on a time-discrete output. After some simplifying assumptions on reconstruction filtering, a time-continuous output can be calculated, arriving at the same result as [16]. Recently, conversion matrices were also proposed to reproduce and extend some earlier N-path analysis results [38]. Still, all these analysis procedures are rather cumbersome, so that it may be beneficial to make a priori assumptions that simplify analysis, while still providing a good approximation, albeit with a more limited scope.

Fig. 5. a) 8-path mixer-first receiver [28, 29]; b) in-band LTI model.

The transfer function and input impedance of the circuit in Fig. 2a for a sinewave voltage input can be approximated assuming $RC \gg T_{on}$ and a virtually constant voltage on the capacitors [28-31, 39]. The difference between the sinewave input V_{in} and the constant capacitor voltage V_{out} is now across resistor R, and the average power dissipation can be evaluated. This leads to a reduction in amplitude related to $\text{sinc}(\pi/N)$. In [28, 29] this dissipation is modelled in an LTI baseband model by an equivalent impedance Z_{sh} (see Fig. 5b). Baseband resistance in parallel to C, e.g. $R_{fb}/(1+A_v)$ in Fig. 5a, resulting in an extra loss resistance. This LTI model is extended in [39] to derive design equations for important properties of a mixer-first receiver, and the relation to the number of paths/phases.

Another analysis method assumes a current driven mixer with 25% duty cycle driving four baseband lowpass impedances $Z_{BB}(f)$ [20, 40]. The RF source is now modelled as a current source with parallel impedance $Z_s(f_{RF})$. It is assumed that the RF-input frequency f_{RF} is confined to $f_s/2$ and $3f_s/2$, and the current entering the switched 4-path filter is approximated by a Fourier series. An approximation of the RF input voltage in the form of an infinite sum is found where not only Z_s around $f_{RF} \approx f_s$ matters, but also around $f_{RF} \approx (4k+1)f_s$. In other words: the source impedance Z_s “seen by the switched mixer” at harmonics of the switching frequency is important. Also, imperfections like clock mismatches and phase noise are analyzed in [20]. However, due to the input frequency limitation, signal folding from frequencies higher than $3f_s/2$ cannot be analyzed in contrast to [16].

III. MIXER-FIRST EVOLUTION

In contrast to traditional radio receivers, mixer-first receivers lack a Low Noise Amplifier (LNA) before the mixer. One of the early mixer-first receivers mainly removes the LNA to save power [26]. The quadrature receiver with 25% duty cycle achieves <6dB Noise Figure and IIP3=-7.5dBm at only 720μW power consumption.

Another motivation to remove the LNA is spurious free dynamic range (SFDR), which is determined by both noise figure and IIP3 [27]. As receiver noise figure is often only a

Fig. 6. N-Path filtering enhanced by discrete time switched-capacitor charge-domain 3rd order Infinite Impulse Response ("3IIR") filtering [41].

few dB above the fundamental 0dB limit, there is more room at the top of the dynamic range by improving IIP3. As passive mixers typically have >20dB better IIP3 than LNAs, while op-amps can realize high linearity in baseband, it makes sense to remove the LNA [27]. An in-band linearity of +11dBm was achieved with NF<6.5dB at 65mW rendering SFDR>79dB in 1MHz bandwidth, which compares favorably in SFDR to professional spectrum analyzers.

Probably the most influential papers on mixer-first receivers are from Molnar's group [28, 29], presenting several useful concepts (see Fig. 5a):

1) Input impedance matching is provided combining switched-C loss, switch resistance and the input impedance of the baseband amplifier, allowing for a NF only slightly higher than 3dB.

2) 8-phase mixing, instead of 4-phase is used so that the filtered staircase approximation of the RF signal contains fewer harmonics. This lowers NF, harmonic re-radiation and harmonic down-conversion (see also [30, 39]). LO-radiation reduction by calibration is discussed in [42, 43].

3) The baseband amplifiers contain tunable feedback resistors connected between the I and Q channels (not shown in Fig. 5a). With this complex feedback, a non-real input impedance can be realized at the RF port, allowing for matching to complex antenna impedances and compensation of input parasitics, resulting in an S_{11} notch that can be tuned to the LO frequency [28, 30].

With the techniques discussed above, IIP3>25dBm and blocker compression B1dB>0dBm is feasible for high offset frequencies. Still, SAW filters can achieve significantly higher values at smaller offset frequencies. Hence, alternative techniques to improve linearity and blocker filtering further have been explored. One example combines N-path filtering with switched-capacitor Infinite Impulse Response (IIR) charge domain filtering (see Fig. 6). The higher-order IIR-filtering renders steeper filtering and better blocker rejection rendering a B1dB≈15dBm at 30MHz offset, albeit at 7-10dB Noise figure [41]).

Another passive pre-filter approach targeting extreme linearity exploits cross-coupling of a switched-R mixer with a switch-R-C N-path notch filter as shown in Fig. 7 [44]. Close to the switching frequency, the notch-filter is high-ohmic and signal is passed. However, the cross-coupled path is low-ohmic for out-of-band blockers, circulating blocker current back to the RF-source (see Fig. 7), so that it does not "reach" the TIA, leading to an IIP3 as high as +42dBm [44].

Fig. 7. Switched-R mixer with differential RF-input, with a cross-coupled switch-RC path for blocker by-passing. Due to the symmetry and high-linearity, an IIP3>+42dBm and B1dB>11dBm is achieved [44].

Mixer-first receivers often realize very good linearity, but usually at a noise figure in the range of 3-6dB for good impedance matching. However, it is possible to go below 3dB adding controlled positive feedback to help realize matching [45]. Also, combining low switch resistance with careful trans-impedance amplifier design, 2.6dB Noise figure has been demonstrated [43].

IV. N-PATH FILTER EVOLUTION

One of the early papers proposing >100MHz RF-frequency N-path filters with a time-continuous output is [15]. It demonstrates a single-ended 8-path bandpass filter driven by a CMOS ring oscillator, with high quality factor of up to 300, while realizing 240-530 MHz tuning range. IIP3 and out-of-band blocking results are not reported.

In our group we (re-)invented the switched-capacitor band-pass filters and only discovered older work during publication attempts. Triggered by good linearity results of [27], we explored a fully differential implemented 4-path filter, focusing on programmability and linearity. A tuning range of 100MHz – 1.2GHz was demonstrated with IIP3>18dBm and 2dBm 1dB-compression point [23]. Notch filters, both with a differential and single input, were also evaluated in 65nm and analyzed in-depth [23].

Mirzaei et al exploited N-path filters in [20, 46] to realize a quad-band SAW-less receiver exploiting to meet all GSM/EDGE requirements, in particular the 15dB Blocker noise figure requirements for a 0dBm blocker at 80MHz offset frequency. It achieves >+2dBm out-of-band 1dB blocker compression. As discussed in the previous section, the work in [20, 46] also greatly contributed to analysis.

In [19] the use of N-path filters to realize filtering in a superheterodyne receiver architecture is explored, with a passband at the difference of two clock frequencies. Using 110MHz IF-frequency, 2.8dB Noise Figure and -8dBm IIP3 is achieved. Image rejection and harmonic rejection is crucial in heterodyne receivers, as explored further in [47]. A flexible Weaver architecture augmented with harmonic rejection was proposed, with a tunable IF-frequency between 100 and 500MHz. As image rejection and harmonic rejection for some conversions is limited to 20-45dB, additional pre-filtering is required for most practical applications.

Ideally, capacitors shunt all out-of-band signal to ground. However, switch resistance limits the conductance to ground,

Fig. 8. Improving filter shape by transconductor coupling between nodes with quadrature signals shift [24].

and hence filter attenuation. The achievable attenuation can be estimated with a simple resistive divider model as shown in Fig. 3. Using a second set of switches that sense the capacitor voltages can largely eliminate the problem [22, 47].

Another limitation of N-path filters is their “rounded filter shape” associated with the real poles. This can be improved as shown in Fig. 8, if two N-path filters with shifted center frequencies are used [22]. This shift is implemented by transconductors connected between baseband nodes with 90 degrees phase shift, contributing a g_m/C shift in frequency (see Fig. 8). However, baseband transconductors add $1/f$ noise, while this problem does not exist if transconductors are used at RF. To approximate the ideal brick-wall filter shape closer, it would be desired to couple multiple stages of high-Q resonators. This cannot be done directly with switched capacitor N-path filter sections, as putting them in parallel simply results in a second order R-L-C with reduced bandwidth (more C). As suggested by Heinlein [48], it is possible to couple N-path band-pass filter sections by gyrators. Fig. 9 shows the block schematic of an 8-path 6-order band-pass filter [24], with a fully passive first section for high linearity. A nice flat-top band-pass filter shape is realized, with 19dB gain and 3dB Noise Figure. The filter has 8MHz bandwidth and shows an out-of-band IIP3 of +26dBm and compression point of +7dBm at 50MHz frequency offset, tunable from 100MHz to 1.2GHz.

In order to increase flexibility for software defined radio, reconfigurable N-path filters have also been explored. In [49] both reconfigurable band-pass and band-reject filtering is demonstrated, while the gm-C shift technique from Fig. 8 is used to allow for continuous shifts of the N-path center-frequency with up to 100MHz.

Other work focusses on improving N-path filter loss and out-of-band rejection, by putting N-path filter sections between inductors, like in a synthetic transmission-line [50]. This tunes out capacitive parasitics, reducing loss and noise, while also reducing folding (see [37]). Other work in advanced SOI technology demonstrates improved linearity and increased frequency range [51]. Moreover, [52] demonstrates that harmonic folding can be limited to <-60dBc using a 16-path band-pass filter.

Fig. 9. Sixth-order band-pass-filter-LNA with one passive 8-path BPF-section and two others coupled by a transconductor and gyrator [24].

V. REDUCING POWER CONSUMPTION

N-path filter bandwidth is inversely proportional to the NRC product (see eqn.(1)), so if the filter is operating in a 50ohm environment a big total capacitance is required to realize a narrow bandwidth (e.g. 32nF for 200kHz GSM-bandwidth [54]). Switch resistance should be much smaller than source resistance R to minimize noise and achieve good filter attenuation (see Fig. 3). However, lowering switch resistance requires wider transistors with larger gate capacitance and hence more dynamic power consumption.

Fig. 10. Increasing the N-path impedance level by: a) Exploiting a Transconductor with load resistance $R_o > R$ [53]; b) Exploiting the Miller effect (“Miller Band-Pass Filter” or “Gain-boosted N-path filter”) [36, 54].

Hence attempts have been made to increase the impedance level by combining an N-path filter with an LNA or LNTA, as shown in Fig. 10. In [53, 55, 56] an LNTA which can be modelled as a transconductance G_m and output resistance R_o precedes the N-path filter, where mainly output resistance R_o defines the R seen by the switched-capacitor N-path filter as shown Fig. 10a (assuming R_f is high for self-biasing). If down-conversion is wanted anyway, the baseband voltages can directly be used as output, while the LNA output voltage is filtered by N-path filtering, improving interference robustness. The output resistance R_o of the LNTA can be significantly higher than 50ohm. If it is e.g. 10x higher, this leads to 10x lower capacitance for the same bandwidth, 10x higher allowed switch-resistance and hence 10x lower dynamic power

Fig. 11. LNA with N-path filter in the feedback path consisting of a series capacitor realizing a notch filter [23], and grounded shunt capacitors that realize the attenuation notches in the overall transfer function [60].

consumption. As blockers are now amplified in the LNA despite of the N-path filtering, a high supply voltage may be desired in the LNA to allow for enough voltage swing without clipping [53]. Another motivation for this LNA-first architecture followed by N-path filtering is lower noise figure [57]. To maximize the output resistance and still cope with strong out-of-band blockers, an LNTA can also be used as has been proposed in [55, 56]. Now the parasitic output resistance of the LNTA acts as resistor R for the N-path filter and as an N-path filter is high-ohmic in-band, there will be significant voltage gain in-band [55, 56]. The LNTA actually works as LNA with voltage gain in-band, but the voltage gain quickly drops out-of-band where the N-path filter is low-ohmic, and the desired LNTA behavior as linear V-I converter is realized (low output voltage swing). Hence improved linearity and compression are still possible at sufficient frequency offset from the N-path filter center frequency. On the other hand linearity is significantly worse than in passive-first receivers, although hybrid approaches are also possible, see e.g. [25, 58]. Overall, significant power consumption savings are certainly possible, as the N-path filter switches can be much smaller than in a 50ohm environment, see for instance the power comparison in [59].

An alternative approach is shown in Fig. 10b, known as “gain-boosted N-path filter” [36] or as “Miller Bandpass Filter” [54]. Here an N-path notch-filter [17] is put in the feedback path of a voltage amplifying LNA, consisting of transconductance G_m and load resistance R_L . The lack of feedback at the notch-frequency leads to a peak-gain of the total negative feedback circuit, i.e. an LNA with band-pass filtering. The input signal V_i is filtered now by an N-path filter, but in a “gain-boosted” way or benefiting from the Miller effect, which has two key advantages:

- 1) Due to the Miller effect, the feedback capacitors see (A_v+1) times more voltage than the input voltage and the N-path filter impedance seen from the input is lowered by this factor. Hence (A_v+1) times less capacitance is needed.
- 2) For the same N-path filter attenuation and noise, the on-resistance of the switches can be increased by a factor

Fig. 12. Gain-Boosted 4-Path with including downconversion mixing [61].

(A_v+1) , reducing switch size and saving considerable dynamic power consumption in the drivers of the switch transistors.

This concept of a BPF-LNA, due to a notch filter in the feedback, can be enhanced further by using the capacitive π -network shown in Fig. 11 [60]. This introduces extra notches useful for TX-leakage suppression. Combining this technique with large overdrive voltage for the switches using thick-oxide transistors in 32nm SOI CMOS technology, an out-of-band IIP3 of +36dBm and B1dB of +17dBm is achieved.

An interesting recent idea that targets low power combines the gain boosted N-path filtering in the LNA with frequency down-conversion as shown in Fig. 12. As the switched-C notch-filter capacitors contain baseband information, this can be sensed differentially and low-pass filtered without additional switches, leading to a compact low power receiver [61]. Other papers target ultra-low-power N-path filters by increasing impedance level and using only three clock phases [62] or even only two phases, if phase information and image rejection is irrelevant [63].

VI. DISCUSSION AND NEW APPLICATION DIRECTIONS

As discussed above, N-path filtering can bring flexibility and high linearity. To benchmark the achieved results, Fig. 13 shows out-of-band IIP3 versus offset frequency normalized to -3dB bandwidth. Not surprisingly, the best linearity comes at a price in Noise Figure (see labels in Fig. 13). As might be expected, the compact power-efficient concepts Fig. 10 achieve modest linearity (~ 10 dBm) because the non-linear properties of transconductors degrade the linearity performance compared to what is possible with passive switch-R-C circuits. As different applications require different performance, there seems place for multiple concepts.

Apart from software defined radio applications, other applications are also explored. Below we name a few. N-path filters can not only be used for frequency-domain filtering, but also for spatial domain filtering, i.e. beamforming, as demonstrated in [55, 59, 64–66]. Especially if implemented at the first node of the receiver, very high compression point and

Fig. 13. Comparison of IIP3 versus offset frequency normalized to -3dB bandwidth for different publications. NF and test frequency are indicated.

IIP3 is possible [66]. Another interesting property of N-path filtering is the controllable gain and group delay [67], which was exploited to improve wideband self-interference cancellation for an in-channel full duplex transceiver [68]. Intriguingly, N-path filters can show non-reciprocal behavior allowing the realization of circulator-like behavior [69].

VII. CONCLUSIONS

During the last decade N-path filtering and mixer-first receivers attracted a lot of research interest, as highly linear high-Q filtering is feasible around a frequency that is digitally accurately controllable. Very promising performance has been demonstrated in the low GHz range, e.g. blocker compression points $>+10\text{dBm}$, Out-of-band IIP3 $>40\text{dBm}$ and a 0dBm blocker noise figure $<5\text{dB}$. Key challenges that remain are the realization of low phase noise multi-phase clocks at acceptable power consumption, and the implementation of sufficient pre-filtering to limit folding adequately.

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